RESEARCH DEPARTMENT

# Class-AB amplifiers employing a complementary pair of transistors:

## improved method for negative feedback

TECHNOLOGICAL REPORT No. G-098

1965/37

THE BRITISH BROADCASTING CORPORATION
ENGINEERING DIVISION

RESEARCH DEPARTMENT

# CLASS-AB AMPLIFIERS EMPLOYING A COMPLEMENTARY PAIR OF TRANSISTORS: IMPROVED METHOD FOR NEGATIVE FEEDBACK

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(1965/37)

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## SUMMARY

This note describes a method of applying current negative feedback to transformerless audio class-AB transistor stages. The output impedance of the circuit is very high but the circuit may also be conveniently adapted to provide a matched output resistance, for example, for a 600-ohm line.

## 1. INTRODUCTION

Circuits employing p-n-p and n-p-n transistors (as complementary pairs) are in widespread use as transformerless audio class-AB output stages, the most common circuit being the emitter-follower circuit shown in Fig. 1. The output impedance is inherently low and is further reduced if overall feedback is applied from point X in Fig. 1.

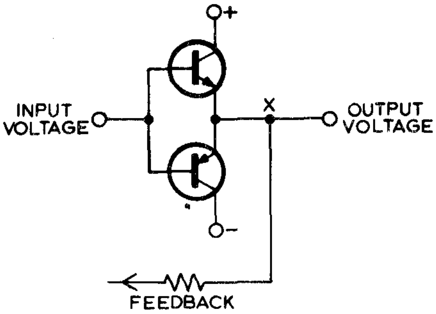

*Fig. 1 - Typical class-AB transformerless output stage*

When a high output impedance is required, however, it is not sufficient to interchange the positions of the transistors (so that the output is taken from the collectors) because the overall feedback maintains the output impedance at a relatively low level.

Another method of obtaining overall feedback employs the voltage across a very small resistor in series with the load, the feedback then being proportional to the load current rather than the load voltage. This raises the output impedance, but has the disadvantage that, either the load or the feedback resistor is not grounded at one end. The circuit described in the next section overcomes this disadvantage.

## 2. PRINCIPLE OF OPERATION OF NEW CIRCUIT

One simple way of applying current negative feedback to an output stage is to put a resistor in each emitter, thus providing local feedback, as shown in Fig. 2. The circuit is then substantially symmetrical despite variations between the transistors and the output impedance is high. It does, however, suffer from one serious disadvantage: when both of the transistors are conducting, the conversion conductance is $2/R_E$ but when only one transistor is conducting it is $1/R_E$. This can be overcome by pure class-B operation but this is not a practical solution.

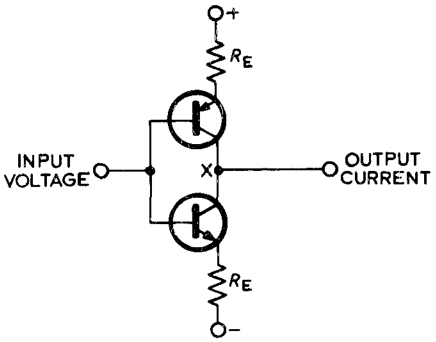

*Fig. 2 - Transformerless class-AB output stage providing a high output impedance*

Overall feedback from point X in Fig. 2 is not suitable - as noted above - but the difficulty may be overcome by taking feedback voltages from both sides of the circuit as shown in Fig. 3. This provides overall current feedback because, neglecting base currents, all the load current must flow through the emitter resistors $R_E$. With this circuit and sufficient loop gain, the overall current gain is $(R_F + R_E)/R_E$ irrespective of whether one or both transistors are conducting. In practical terms, a loop gain of about 10 is great enough (that is $R_T \geqslant 10(R_F + R_E)$ where $R_T$ is the transfer resistance of the drive amplifier).

Thus, the circuit has an output impedance which is very high (over 1 MΩ with low-power circuits), a well defined gain and a relatively large bandwidth. There is no cross-over distortion and the total distortion, even at full output, can be kept small.

There are two disadvantages: first, the circuit is not as efficient as the emitter-follower output circuit because of the power developed in the emitter resistors, and second, the feedback resistances must be matched. Further consideration of this aspect and the degree of matching required in the transistors is given in section 3.2.

One possible application of the circuit is shown in Fig. 4 where the resistance $R_0$ within the amplifier provides a match to the load. With suitable component values it is possible to develop a peak-to-peak output voltage swing which is about 90% of the supply voltage as opposed to just under 50% for the case of a voltage output and series matching resistor.

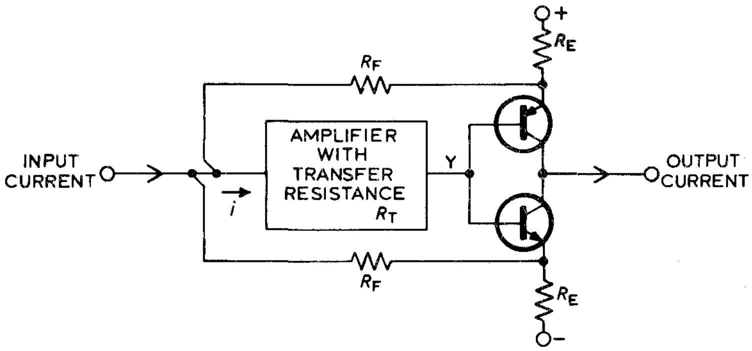

*Fig. 3 - The circuit in Fig. 2 with overall feedback*

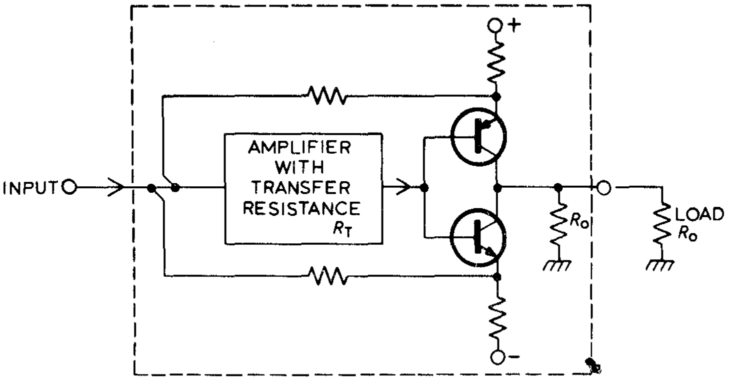

*Fig. 4 - An output stage with matched output resistance*

# 3. DESIGN AND PERFORMANCE

## 3.1. Design

The design of the amplifier follows the simple principles given by Cherry and Hooper[1] for class A circuits. It will generally be found convenient to start by choosing a suitable value for $R_E$. A large value gives good bandwidth and low distortion while a low value gives good efficiency and high gain. Where a large output voltage is required, $R_E$ should be about one tenth of the effective load resistance.

Having chosen $R_E$, the quiescent output current $I_Q$ is chosen to give about 14 dB of local feedback: thus, for $R_E = 5r_e$, where $r_e = 25/I_Q$,

$$I_Q \approx \frac{125}{R_E} \text{ mA}$$

Depending on the form and gain of the rest of the circuit, $R_F$ is chosen to give the required degree of overall feedback.

## 3.2. Performance

As already noted, the current gain of the circuit (Fig. 3) is approximately $(R_F + R_E)/R_E$. The bandwidth at low output levels is determined mainly by the capacitance due to both the Miller effect and the base-emitter capacitances which appears at point Y (Fig. 3). The overall feedback extends the bandwidth at low levels but at high levels it is necessary to ensure that the drive amplifier can develop the required voltage across this capacitance and also that the transistors will turn on (and off) sufficiently quickly.

Distortion at low levels, that is when both transistors are conducting, is low firstly because of the negative feedback and secondly because there is second-harmonic cancellation between the two halves. When the output is such that each transistor just cuts off in one direction the ratio of the amplitude of the second harmonic to that of the fundamental is approximately[2]:

$$\frac{1}{4}\frac{r_e}{R_E}\frac{R_F}{R_T}K$$

where K is the degree of cancellation between the halves and $r_e = 25/I_Q$. Using the minimum values suggested above ($R_E = 5r_e = 125/I_Q$ and $R_T = 10\ R_F$) and assuming 14 dB of cancellation the second-harmonic amplitude is 0·1% of that of the fundamental.

When the peak output current swing is twice the quiescent current there is a maximum proportion of third harmonic given approximately by:

$$\frac{1}{12}\frac{R_F}{R_T}$$

This expression depends only on the loop gain and has a value of 0·82% when the loop gain ($R_T/R_F$) is 10.

At full output the distortion is determined by the degree of matching between the feedback resistors and between the transistors. It should be noted that the feedback controls the emitter current and that the collector current is very slightly different from this. Thus, if the current gains of the transistors are $\beta_1$ and $\beta_2$, the proportion of second harmonic is:

$$\frac{1}{2}\left(\frac{1}{\beta_1} - \frac{1}{\beta_2}\right)$$

For low distortion the transistors must either have high gain or be reasonably well matched. For example, in a typical circuit one transistor might have a gain of 40 and the other a gain of 120. The second harmonic due to this cause would then be 0·8% at full output.

The four feedback resistors must likewise be matched so that the ratio $(R_F + R_E)/R_E$ on one side is equal to $(R_F + R_E)/R_E$ on the other side. Using ± 2% tolerance resistors it is possible that there would be 4% second harmonic at full output. This situation is most unlikely, however, because all the resistors would have to be at the extremes of their tolerance and in a particular combination.

Where the very lowest distortion is required, the two halves of the amplifier can be balanced by the use of a small potentiometer as shown in Fig. 5.

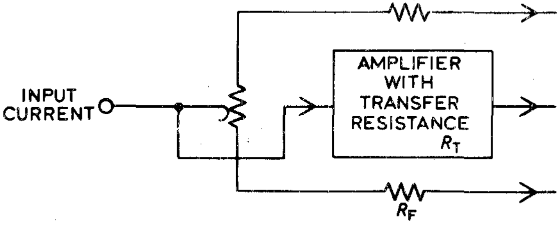

*Fig. 5 - A possible method of reducing second-harmonic distortion*

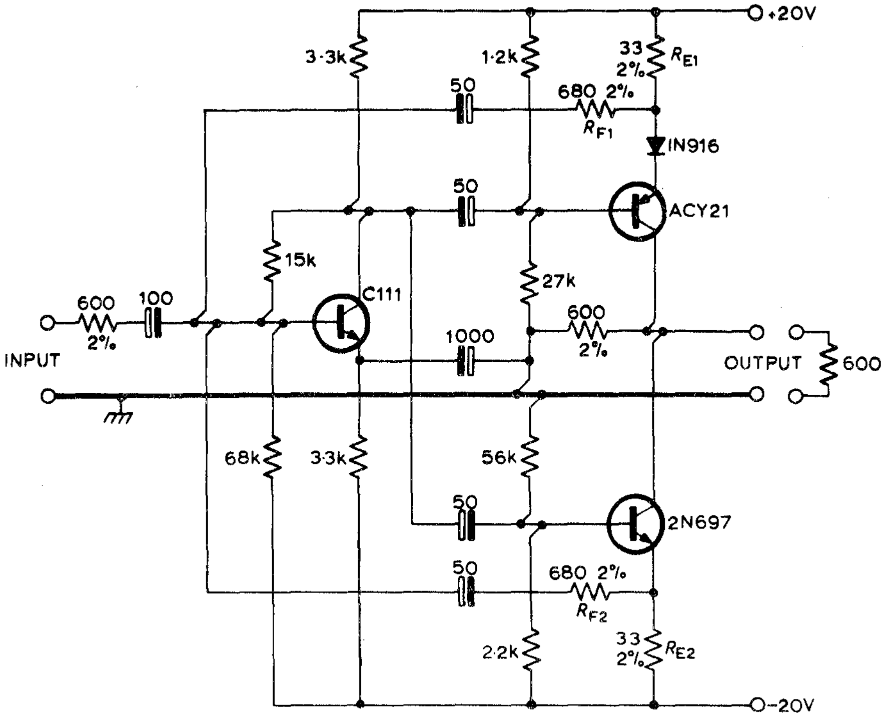

*Fig. 6 - A practical amplifier*

## 4. A PRACTICAL EXAMPLE

Fig. 6 shows a practical circuit constructed to test the principles given above. The amplifier is designed to provide a matched output to a 600Ω load and the input impedance is likewise 600Ω. The emitter feedback resistors stabilize the quiescent current as well as providing a.c. feedback and the silicon diode (1 N 916) in series with the germanium transistor (ACY 21) prevents[3] thermal run-away at high outputs. The loop gain is 20 times and the overall gain is 9·3 times (when loaded by 600Ω).

The maximum third-harmonic distortion was 0·3% which is a little less than the computed value of 0·4%. The maximum output voltage was 34 V peak-to-peak; this is 85% of the supply voltage, the maximum supply voltage being limited by the ratings of the ACY 21. The second-harmonic distortion at full output was 0·6%, in reasonable agreement with the value of 0·7% computed from the measured values of the feedback resistors and the current gains of the transistors.* These are shown below:

$R_{F_1} = 687 \cdot 3\Omega$ $\qquad$ $R_{F_2} = 673 \cdot 9\Omega$

$R_{E_1} = 33 \cdot 47\Omega$ $\qquad$ $R_{E_2} = 33 \cdot 12\Omega$

$\beta_1 = 110$ (ACY 21) $\qquad$ $\beta_2 = 70$ (2N697)

At high frequencies the maximum output was limited by the turn-on time of the ACY 21 and was 15 V peak-to-peak at 50 kc/s.

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* The predicted phase of the second harmonic also agreed with the measured phase.

Since completing this work, a design has been published[4] for a tape-recorder amplifier which uses the principle described here to obtain a constant-current drive for the recording head. This resulted in a considerable economy in voltage swing compared with driving the head through a series resistor but, since the quiescent current was many times greater than the peak load current, the application to class-AB amplifiers was not explored.

## 5. CONCLUSIONS

If current negative-feedback is applied to a complementary-pair amplifier in the manner described, the output impedance is high but may be shunted by a resistor to provide a matched output impedance. A 600-ohm audio amplifier was constructed which has successfully demonstrated the principle.

The two main disadvantages are that the circuit is somewhat less efficient than an ideal class-B output stage and two pairs of matched resistors are required.

## 6. REFERENCES

1. CHERRY, E.M. and HOOPER, D.E.: 'The Design of Wide-band Transistor Feedback Amplifiers', Proc.Inst.Elect.Engrs., 1963, **110**, 2, pp. 375 - 389.
2. IZATT, J.B.: 'Distortion in Class A Transistor Amplifier Circuits', Electronics Letters, 1965, **1**, p. 126.
3. DAUGHERTY, D.G. and GRIENER, R.A.: 'Low Distortion and High Thermal Stability in Transistor Audio Power Amplifiers', I.E.E.E. Trans.Audio, 1964, **AU-12**, pp. 26 - 29.
4. WATSON, J.B.: Wireless World, 1965, **71**, 8, p. 403 (letter).

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Printed by BBC Research Department, Kingswood Warren, Tadworth, Surrey